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Method for channel estimate with transmit beamforming

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FIELD OF THE INVENTION

The present invention relates to a method for channel estimation according to the preamble of claim 1.

Such a method may be used in particular in any communication system using UMTS standard.

BACKGROUND OF THE INVENTION

A UMTS communication system, defined in the UMTS standard, comprises at least a base station comprising a transmitter that sends a signal to users, that is to say to receivers through downlink communication channels, and receivers that send a signal to said transmitter of said base station through uplink communication channels.

A downlink communication channel uses two logical channels, a first channel called common pilot channel (CPICH) that continuously transmits known pilot symbols in particular for controlling the transmission of a signal and a second channel called dedicated physical channel (DPCH). Said second channel DPCH comprises two dedicated channels, a first sub-channel called dedicated physical data channel (DPDCH) time multiplex with a second sub-channel called dedicated physical control channel (DPCCH).

Transmit beamforming can be employed at the base station to improve the second logical channel DPCH transmission. It consists in providing dedicated beams for useful data users. It improves the capacity of the UMTS communication system to manage a certain number of users by reducing interferences between two users. On the contrary, beamforming is not suited for the first logical channel CPICH transmission that is to be uniformly broadcast in time and space to all users. As seen from the receiver, the second logical channel DPCH is in general different from the first one CPICH. Usually, when no transmit beamforming is used, in order to optimize the receiver, one uses a maximum a posteriori probability (MAP) optimization, i.e. one uses channel estimates of the CPICH channel as channel estimates for the DPDCH channel. But, when transmit beamforming is applied, channel estimates obtained on the CPICH cannot be directly applied as channel estimates for the DPCH as it is usually done because they both differ.

SUMMARY OF THE INVENTION

Accordingly, it is an object of the invention to provide a method for channel estimation with transmit beamforming, said channel comprising a first and second channels, which achieves an efficient optimization of a receiver without too much complexity.

To this end, there is provided a method as claimed in claim 1.

As we will see in detail further on, such a method enables to optimize a receiver, i.e. reducing interferences between users, by calculating an estimate of the first physical channel from the second physical estimate and moreover from a first logical channel estimate. The use of a first logical channel estimate allows a better tracking of the channels variation in time, said first logical channel transmitting continuously contrary to the second physical channel.

BRIEF DESCRIPTION OF THE DRAWINGS

The invention and additional features, which may be optionally used to implement the invention to advantage, are apparent from and elucidated with reference to the drawings described hereinafter.

- Fig. 1 represents a diagram illustrating the main steps of the method for channel estimation according to the invention, and
- Fig. 2 illustrates an estimator using the method for channel estimation according to the invention.

DETAILED DESCRIPTION OF THE EMBODIMENTS

In the following description, well-known functions or constructions by a person of ordinary skilled in the art are not described in detail since they would obscure the invention in unnecessary detail.

The present invention relates to a method for channel estimation with transmit beamforming, said channel comprising a first CPICH and second DPCH channels. Said method is used in particular in a RAKE receiver of a UMTS communication system, for example in a receiver of a mobile terminal MT.

A UMTS communication system comprises at least a base station BS comprising a transmitter that transmits a signal to users, that is to say to some receivers through downlink communication channels, said receivers sending a signal to said transmitter of said base station BS through uplink communication channels. The transmitted signal comprises useful data and pilot symbols.

A downlink communication channel uses two logical channels, the first one is called common pilot control channel CPICH that continuously transmits known pilot symbols and that employs a fixed spreading factor M equal to 256, and the second one is called dedicated

physical channel DPCH. Said second channel DPCH comprises with useful data user forming a first sub-channel called dedicated physical data channel DPDCH which are time multiplex within a same time slot with known pilot symbols forming a second sub-channel called dedicated physical control channel DPCCH.

Transmit beamforming is employed at the base station BS in order to improve the DPCH reception at the RAKE receiver of a mobile terminal MT. It consists in providing dedicated beams for useful data. It improves the capacity of the UMTS communication system to manage a certain number of users by reducing interferences between two users.

A RAKE receiver comprises fingers that are associated with each path of a beam dedicated to useful data that has enough power, a finger being a filter adapted to recover a path.

It will be noted that logical channels use physical channels in order to transmit their data. Classically a physical channel has a temporal structure and a spatial structure, a physical channel being for example an antenna.

It will be also noted that the transmitted signal is composed with bits, said bits being first encoded, the encoded bits being called symbols. Then, these symbols are modulated (e.g. by a QPSK modulation, well know by the person skilled in the art). Then, said modulation symbols, are submitted to a spreading. The use of a spreading factor M and of an associated code C enable to transmit a signal on a larger bandwidth. After the spreading, the bits are defined in term of a chip. We have the following formula: bit rate * M = chip rate.

In order to recover the useful data that have been received by the UMTS RAKE receiver, one has to estimate the transmission channels such as the downlink user dedicated physical channel DPCH. While such an estimate is done, effects on the received signal due to interferences of such channels can be removed in order to recover said useful data.

As we will see, for such an estimate, there is a first step of model of the physical channels associated said first CPICH and second DPCH channels, said models being a linear superposition of discrete multipath MP components (characterized by the term $C_p(t)\delta(\tau-\tau_p)$, as we will see below) following an uncorrelated-scattering wide-sense stationary model. A multipath component is characterized by a time-varying complex coefficient $c_p(t)$ and a propagation delay τ_p. Said time-varying complex coefficient has a CPICH multipath coefficient $c_{pcpich}(t)$ and a p-th DPCH multipath coefficient

$c_{pdpch}(t) = \beta_p c_{pcpich}(t)$ which depends on a complex factor β_p due to the transmit beamforming of the base station BS.

Signal Model

We first begin to model a general channel h. This general model will be applied to the CPICH and DPCH channels afterwards.

1) General Signal Model

We consider a general case of the transmission of a signal $s(t)$ through a multipath channel h. A continuous-time complex base-band received signal at a RAKE receiver is given by

$$y(t) = x(t) + v(t) \tag{1}$$

$v(t)$ defines noise plus interference,
y(t) is a received signal before despreading and before an A/D conversion,
x(t) is a received signal that comprises the useful data, which have been subjected to a distortion due to the effects of the channel h.

$x(t)$ is given by

$$x(t) = \int h(t, t-\tau) z(\tau) \mathrm{d}\tau \tag{2}$$

where $h(t,\tau)$ is a time-varying multipath channel impulse response and

$$z(t) = \sum_k a(k) \psi(t - kT_c) \tag{3}$$

where:

- z(t) is a signal just after a D/A conversion in the transmitter,
- T_c, is a chip period,
- $a(k) = s(\lfloor k/M \rfloor) * d(k)$, M denotes the spreading factor, and the term $\lfloor\ \rfloor$ is a rounded down,
- $s(n)$ is an n-th modulation symbol (e.g., QPSK),

- $d(k)$ is a k-th chip, and
- $\psi(t)$ is a pulse-shaping filter that limits the base-band of the receiver.

Note that this model can be used by systems with several spreading layers, like DS-CDMA system defined by the standard IS-95.

The channel h is assumed to follow a wide-sense stationary uncorrelated scattering model with Rayleigh fading and multipath response as described in the document hereby incorporated by reference "J. G. Proakis *Digital Communications,* NY: McGraw-Hill, 3rd edition, 1995" which is incorporated by reference. It means that:

$$h(t,\tau)=\sum_{p=1}^{P} c_p(t)\delta(\tau-\tau_p) \qquad (4)$$

where P denotes a number of paths available for one user, and $c_p(t)$ represents a path time-varying complex coefficient at propagation delay τ_p. The path complex coefficients $c_p(t)$ vary depending on the velocity of the mobile terminal MT and on surrounding scattering objects such as buildings, mountains, cars, trains.... $\sum_{p=1}^{P} |c_p(t)|^2$ corresponds to the energy that has gone through a DPCH channel. These complex coefficients $c_p(t)$ are not known by the receiver, because the physical channel characteristics such as the complex coefficients are unknown from said receiver. Therefore, they have to be estimated as we will see later on.

In the case of transmit beamforming through a Q-antennas array of a RAKE receiver, Q being the number of antennas of the array also called rake, the channel model expressed by (4) becomes

$$h(t,\tau)=\sum_{p=1}^{P} w^H a(\theta_p)c_p(t)\delta(\tau-\tau_p) \qquad (5)$$

where:

- $w=[w_0,\ldots,w_{Q-1}]^T$ denotes a transmit beamforming weight vector, which enables to weight the antennas when forming the beams,
- $a(\theta)=[a_0(\theta),\ldots,a_{Q-1}(\theta)]^T$ denotes an antenna array response vector in a direction θ, and

- The superscripts $(\,.\,)^T$ and $(\,.\,)^H$ denotes transpose and Hermitian transpose respectively.

By inspecting equations (4) and (5) one may notice that, during a time slot, transmit beamforming only affects the amplitude of the path time-varying complex coefficients $c_p(t)$ by a complex factor $\mathbf{w}^H\mathbf{a}(\theta_p)$ and not the delays τ_p. It means that the channel's temporal structure is not affected by the antenna, neither by the transmit beamforming.

Moreover the path time-varying complex coefficient $c_p(t)$ stays the same in both case (3) and (4) up to a complex factor $\beta_p = w^H a(\theta_p)$.

2) Received Signal Model for DPCH and CPICH

We now apply this general channel model to the DPCH and CPICH channels as follows.

We consider the case where the first CPICH and second DPCH channels are superposed, that is to say only when the useful data are transmitted, knowing that the first CPICH channel continuously transmits pilot symbols while the second DPCH channel only transmits when useful data have to be transmitted.

In that case, the continuous-time complex base-band received signal (1) can be written as

$$y(t) = x_{cpich}(t) + x_{dpch}(t) + v(t) \tag{6}$$

where $x_{cpich}(t)$ and $x_{dpch}(t)$ are the signal components associated with the first CPICH and second DPCH channels respectively given by

$$x_{cpich}(t) = \int h_{cpich}(t, t-\tau) z_{cpich}(\tau)\mathrm{d}\tau$$

$$x_{dpch}(t) = \int h_{dpch}(t, t-\tau) z_{dpch}(\tau)\mathrm{d}\tau$$

where:

$$h_{cpich}(t,\tau) = \sum_{p=1}^{P} c_p(t)\delta(\tau - \tau_p)$$

$$h_{dpch}(t,\tau) = \sum_{p=1}^{P} \mathbf{w}^H \mathbf{a}(\theta_p) c_p(t)\delta(\tau - \tau_p)$$

and

$$z_{cpich}(t) = \sum_k a_{cpich}(k)\psi(t - kT_c)$$

$$z_{dpch}(t) = \sum_k a_{dpch}(k)\psi(t - kT_c)$$

where:

- $a_{cpich}(k) = s_{cpich}(\lfloor k/256 \rfloor) * d_{cpich}(k)$,
- $a_{dpch}(k) = s_{dpch}(\lfloor k/M \rfloor) * d_{dpch}(k)$,
- $s_{cpich}(n)$, $s_{dpch}(n)$ are the n-th modulation symbols, associated with the first CPICH and second DPCH channels respectively, and
- $d_{cpich}(k)$, $d_{dpch}(k)$ are the k-th chips, associated with the first CPICH and second DPCH channels respectively.

It will be noted that the spreading factor M may not be the same for the first CPICH and second DPCH channels. In the UMTS standard the first channel CPICH has a fixed spreading factor equal to 256 whereas the second channel DPCH has a spreading factor that can vary between all powers of 2 between 4 and 512.

Furthermore, one can notice that the CPICH is not modulated, i.e., $s_{cpich}(n)=A(1+j)$ for any n symbol, where A is an amplitude normalization factor. It means that $s_{cpich}(n)$ doesn't vary in time.

We assume now that the RAKE receiver has perfectly recovered the path delays as the RAKE receiver knows the delays τ_p of the paths associated to the CPICH and DPCH channels. The RAKE receiver comprises an analog/digital converter that allows to go from a continuous time (t) to a discrete time (n).

We also assume that the RAKE receiver has already de-spread the received signal $y(t)$ in order to detect the CPICH and the sub-channels DPCCH/DPDCH channels. It means that the RAKE receiver has multiplied the received signal $y(t)$ by the code C associated with the spreading factor M used for the transmission of a signal so as to cancel the effect of the spreading and so as to restore the original narrow bandwidth. During the dispreading, a sampling is done according to a symbol rate 1/T, T representing the symbol period.

For the sake of simplicity, we limit the treatment to a single path/finger case of the receiver. Thus, we will drop the subscript p to denote the time-varying complex coefficient $c_p(t)$ hereinafter also called path fading channel coefficient.

Furthermore, in a first instance, we limit the analysis to the case where the second channel's DPCH spreading factor M is equal to the one of the CPICH. We shall generalize the analysis to other spreading factors successively.

Assuming to model all interferences and especially multiple-access interferences (MAI) and inter-path interferences as if it was additive white Gaussian noise (AWGN), the de-spread signals, also called observations for a RAKE finger related to the CPICH and the DPCH channels are respectively given by

$$y_{cpich}(n) = c(n)s_{cpich}(n) + v_{cpich}(n)$$
$$y_{dpch}(n) = \mathbf{w}^H \mathbf{a}(\theta)c(n)s_{dpch}(n) + v_{dpch}(n) \quad (7)$$

where:

- $v_{dpch}(n)$ and $v_{cpich}(n)$ denote noise-plus interference terms associated with the second DPCH and first CPICH channels respectively, classically assumed to be independent from each other,
- $c(n)$ is the path time-varying complex coefficient,
- $s_{dpch}(n)$ and $s_{cpich}(n)$ denote the pilot symbols of the DPCH and CPICH channels respectively, at a symbol time n.

It will be noted that a MAI is for example interferences between different users or interferences for one user due to the first channel CPICH or due to different data flows dedicated to this user, and that an inter-path interference is due to interferences between path impulse responses at different delays τ.

Under the assumption that the first CPICH and second DPCH channels vary every symbol time n and that the interference and the noise can be modeled as an additive white Gaussian noise, we then make a maximum likelihood (ML) static channel estimates at every symbol time n. This ML can be built as follows:

$$\hat{c}_{cpich}(n) = s^*_{cpich}(n)y_{cpich}(n) = c(n) + v_{cpich}(n)$$
$$\hat{c}_{dpch}(n) = s^*_{dpch}(n)y_{dpch}(n) = \gamma\mathbf{w}^H \mathbf{a}(\theta)c(n) + v_{dpch}(n) \quad (8)$$

where, $v_{cpich}(n)$ and $v_{dpch}(n)$ symbolize the AWGN associated with the first CPICH and second DPCH channels respectively, and where without loss of generality, we assumed the pilot symbols of both channels to be normalized such that $|s_{dpch}(n)| = |s_{cpich}(n)| = 1$.

Note that $s^*(n)$ is a pilot symbol multiplied by its complex conjugated.

Note that the pilot symbols $s_{cpich}(n)$ and $s_{dpch}(n)$ of both channels CPICH and DPCH are known by the transmitter and RAKE the receiver. γ is a coefficient that symbolizes a difference of energy between the pilot symbols of the second sub-channel DPCCH channel and the useful data of the first sub-channel DPDCH channel.

Upon the models of the first DPCH and second CPICH channels, the following remarks can be made.

1. Such ML static channel estimates are formed by demodulating

(multiplication by the complex conjugated) observations y_{cpich} and y_{dpch} at the output of a finger in data aided (DA). But it can be done for decision direct (DD) fashion. While DA demodulation is achieved using the known pilot symbols provided on the second sub-channel DPCCH and first logical CPICH channels, DD demodulation requires the use of hard decisions taken at the RAKE receiver output as pilot symbols, when known pilot symbols are not available, as it occurs during the first sub-channel DPDCH channel transmission. It means that when receiving some bits of a signal, a hard decision is taken in order to define a symbol. Such symbol is taken as a pilot symbol.

Therefore, the formula (8) can be written as:

$$s^{*}_{dpch}(n)y_{dpch}(n) = s^{*}_{dpcch}(n)y_{dpcch}(n) + \hat{s}^{*}_{dpdch}(n)y_{dpdch}(n)$$
$$= \mathbf{w}^{H}\mathbf{a}(\theta)c(n) + v_{dpcch}(n) + \gamma\mathbf{w}^{H}\mathbf{a}(\theta)c(n) + v_{dpdch}(n)$$

2. During a symbol rate 1/T, the path time-varying complex coefficients of the first CPICH and second DPCH channels only differ by a complex scalar factor given by $\beta = \gamma\mathbf{w}^{H}\mathbf{a}(\theta)$.

3. When there is a transmission, there is a transmit power offset between both first CPICH and second DPCH channels because the power used to transmit a signal is different for said first CPICH and said second DPCH channels. While the power is relatively constant for the first channel CPICH, the power for the second channel DPCH is changing as a function of some propagation conditions. For example, when the mobile terminal MT is near to the base station BS, the power is lower than when said mobile terminal MT is far from said base station BS. As said power offset between CPICH and DPCH is a real and is constant over a time slot, it can be taken into account in the complex factor β. Hence, DPCH power assignment is included by properly normalizing the term $||w||$.

4. On a physical point of view, to each path on the first channel CPICH corresponds a path on the second channel DPCH with the same delay τ and a same Doppler spread but with different complex paths' amplitudes β and β $c(t)$ and so β $c(n)$, the Doppler spread being related to the Doppler effect made by the mobile terminal's MT movement or made by the scattering objects in movement around the mobile terminal MT. The two complex paths' amplitudes are only different up to a complex factor β that is determined by the scalar product of the beamforming weight vector $\boldsymbol{w}$ and the antenna array response $\mathbf{a}(\theta)$ in correspondence of the direction θ of that path. It means that the two channels CPICH and DPCH behave in the same way up to said complex factor β, which is modified as a function of a path. Hence, these channels CPICH and DPCH are not different in a temporal point of view.

5. On a mathematical point of view, it means that the path time-varying complex coefficients $c(t)$ and so $c(n)$ vary as a function of the time as it has been mentioned before, as said coefficient depends on said Doppler spread.

From these remarks, we can conclude that, as the two channels CPICH and DPCH behave in the same way up to said complex factor β, said complex factor β can be used to estimate more particularly the first physical channel DPDCH. Hence, we have to calculate first an estimate of said complex factor β.

To this end, we use the DPCH pilot symbols jointly with the pilot symbols provided by the CPICH for estimating the unknown complex factor β. Once an estimate $\hat{\beta}$ of said complex factor β is calculated, a channel estimate for the first sub-channel DPDCH channel will be easily obtained by multiplying the estimate $\hat{\beta}$ of said complex factor by the channel estimates obtained over the CPICH channel.

It will be noted that this approach can include the DA ML only or with the DD ML channel estimates according to equation (8) in the estimate of β.

Note that this approach is more advantageous than the one that consists in using only the pilot symbols provided by the second sub-channel DPCCH to directly estimate the first sub-channel DPDCH. Indeed, the number of pilot symbols provided by said second sub-channel DPCCH is usually fairly small (from 1 to a maximum of 16 depending on the slot format), and in some cases not enough to provide sufficient noise suppression for a reliable DPDCH first sub-channel estimate. Moreover, as the second sub-channel DPCCH is not continuously transmitting pilot symbols, at high mobile terminal T velocity, there is a risk of not taking into account some variations of the first sub-channel DPDCH during its transmission while said second sub-channel DPCCH is not transmitted, such that the channel estimates corresponding to said sub-channel DPCCH transmitting period cannot be applied as channel estimates for the whole successive first sub-channel DPDCH period transmission. With the first channel CPICH that transmits continuously some pilot symbols, those drawbacks are overturned.

DPDCH Channel estimate from CPICH and DPCCH Channel estimates

Therefore in order to estimate the first sub-channel DPDCH, we process each multipath component as follows.

1) In a first step, we build DPCH and CPICH instantaneous maximum likelihood ML channel multipath MP coefficients estimates $\hat{c}_{dpch}(n)$, $\hat{c}_{cpich}(n)$.

As above described in the paragraph Received Signal Model for DPCH and CPICH, the estimates are:

$$\hat{c}_{cpich}(n) = s^{*}_{cpich}(n) y_{cpich}(n) = c(n) + v_{cpich}(n)$$
$$\hat{c}_{dpch}(n) = s^{*}_{dpch}(n) y_{dpch}(n) = \gamma \mathbf{w}^{H} \mathbf{a}(\theta) c(n) + v_{dpch}(n)$$

In a preferred embodiment, we use the DA fashion: The DPCH instantaneous maximum likelihood ML channel multipath MP coefficient estimate $\hat{c}_{dpch}(n)$ is only based on the ML coefficient estimate of the second sub-channel DPCCH. That is to say, only the estimate of the second sub-channel DPCCH is used jointly with the estimate of the CPICH channel.

Hence, the estimates are:

$$\hat{c}_{cpich}(n) = s^{*}_{cpich}(n) y_{cpich}(n) = c(n) + v_{cpich}(n)$$
$$\hat{c}_{dpch}(n) = s^{*}_{dpch}(n) y_{dpch}(n) = \mathbf{w}^{H} \mathbf{a}(\theta) c(n) + v_{dpch}(n)$$

2) In a second step, we build an estimate $\tilde{c}_{cpich}(n)$ of the CPICH multipath coefficient.

In a non-limitative preferred embodiment, the CPICH multipath coefficient's estimate is done by a linear prediction as described in the document hereby incorporated by reference [1] "J. Bsltersee, G. Fock, P. Schultz-Rittich, and H. Meyr, "Performance analysis of phasor estimation algorithms for FDD-UMTS RAKE receiver," *IEEE 6th Symp. on Spread Spectrum Techonologies and Applications,* NJIT, NJ, September 2000."

Under the assumption that the channel paths are Gaussian complex random process, a maximum a posteriori probability (MAP) channel path coefficient estimate is given by a conditional expected $\hat{c}_{MAP}(n)$ value given the whole observation ***y(n)***

$\mathbf{y}(n) = [y_{cpich}(n - K_{cpich} + 1) \quad \ldots \quad y_{cpich}(n) \quad]^{T}$ and the corresponding pilot symbols sequence $\mathbf{s}(n) = [s_{cpich}(n - K_{cpich} + 1) \quad \ldots \quad s_{cpich}(n) \quad]^{T}$ such as

$$\hat{c}_{MAP}(n) = E\{\hat{c}_{ML}(n) \mid y(n), s(n)\} \qquad (9)$$

where $\hat{c}_{ML}(n)$ represents the instantaneous ML static channel estimate, K_{cpich} is the number of available pilot symbols from the CPICH channel.

In the document referenced [1], it is shown that a MAP optimal channel estimate (9) can be built by low-pass filtering instantaneous ML channel estimates (8), i.e.,

$$\hat{c}_{MAP}(n) = f^{H}\hat{c}_{ML}(n) \qquad (10)$$

where $\hat{\mathbf{c}}_{ML}(n) = [\hat{c}_{cpich}(n) \quad \dots \quad \hat{c}_{cpich}(n-K_{cpich}+1) \quad]^{T}$ denotes the vector of the ML channel estimates and $f = [f(N-1) \quad \dots \quad f(0)]^{T}$, with $N = K_{cpich}$, denotes the low pass filter coefficients respectively.

Although the current formulation corresponds to a classical forward linear prediction (we go from n to n-K+1...), we shall remark that it can trivially extent to both, backward or forward-backward (i.e., two-sided) linear prediction.

The linear prediction is well known by the person of ordinary skill in the art and therefore won't be further described.

Note that for a DD fashion, we will have the formula (9) extents to the DPCH channel such as:

$$\mathbf{y}(n) = [y_{cpich}(n-K_{cpich}+1) \quad \dots \quad y_{cpich}(n) \quad y_{dpch}(n-K_{dpch}+1) \quad \dots \quad y_{dpch}(n)]^{T}$$

and the corresponding pilot symbols sequence

$\mathbf{s}(n) = [s_{cpich}(n-K_{cpich}+1) \quad \dots \quad s_{cpich}(n) \quad s_{dpch}(n-K_{dpch}+1) \quad \dots \quad s_{dpch}(n)]^{T}$ such as

$$\hat{c}_{MAP}(n) = E\{\hat{c}_{ML}(n) \mid y(n), s(n)\} \qquad (9)$$

where $\hat{c}_{ML}(n)$ represents the instantaneous ML static channel estimate, K_{cpich} and K_{dpch} is the number of available pilot symbols from the CPICH channel and DPCCH sub-channel respectively.

3) In a third step, we build estimates of a DPCH and CPICH cross-correlation $\mathrm{E}\{\hat{c}_{dpch}(n)\hat{c}_{cpich}(n-l)^{*}\}$ and of a CPICH auto-correlation $\mathrm{E}\{\hat{c}_{cpich}(n)\hat{c}_{cpich}(n-l)^{*}\}$ from maximum likelihood ML estimates of step 2 at non-zero lag for noise suppression **($\tau \neq 0$)**.

The estimates are built as follows.

From the formula (8), we have:

$$\hat{c}_{cpich}(n) = c(n) + v_{cpich}(n)$$
$$\hat{c}_{dpch}(n) = \gamma \mathbf{w}^H \mathbf{a}(\theta) c(n) + v_{dpch}(n)$$ with the complex scalar factor given by $\beta = \gamma \mathbf{w}^H \mathbf{a}(\theta)$.

In a preferred embodiment, only the DA fashion is used: the cross-correlation of the DPCH and CPICH channels is a cross-correlation of the second sub-channel DPCCH and first channel CPICH.

Hence, we have:

$$\hat{c}_{cpich}(n) = c(n) + v_{cpich}(n)$$
$$\hat{c}_{dpch}(n) = \mathbf{w}^H \mathbf{a}(\theta) c(n) + v_{dpch}(n)$$ with the complex scalar factor given by $\beta = \mathbf{w}^H \mathbf{a}(\theta)$.

The cross-correlation $\phi_{dc}(k)$ of the DPCCH and CPICH ML channel estimates, at lag l, is:

$$\phi_{dc}(l) = \mathrm{E}\{\hat{c}_{dpch}(n)\hat{c}_{cpich}(n-l)^*\}$$
$$= \mathrm{E}\{(\beta \quad c(n) + v_{dpch}(n))(\hat{c} \quad (n-l)^* + v_{cpich}(n-l))\} =$$
$$\mathrm{E}\{\beta \quad c(n)\hat{c}(n-l)^* + v_{dpch}(n) \bullet \hat{c}(n-l)^* + \beta \quad c(n) \bullet v_{cpich}(n-l) + v_{dpch}(n) \bullet v_{cpich}(n-l)\}$$

One can notice that when we take the AWGN model, noise samples $v_{cpich}(n)$ and $v_{dpch}(n)$ taken at any lag $l \neq 0$ are uncorrelated. Therefore, their means are null. Hence,

$$\phi_{dc}(l) = \mathrm{E}\{\beta c(n)\hat{c}(n-l)^*\}$$
$$= \beta \ \mathrm{E}\{c(n)\hat{c}(n-l)^*\}$$

Thus, $\beta = \phi_{dc}(l) / \mathrm{E}\{c(n)\hat{c}(n-l)^*\}$

The auto-correlation $\phi_{cc}(l)$ of the CPICH channel estimate, at lag l, is:

$$\phi_{cc}(l) = \mathrm{E}\{\hat{c}_{cpich}(n)\hat{c}_{cpich}(n-l)^*\}$$
$$= \mathrm{E}\{(c(n) + v_{cpich}(n))(\hat{c} \quad (n-l)^* + v_{cpich}(n-l))\} =$$
$$\mathrm{E}\{c(n)\hat{c}(n-l)^* + v_{cpich}(n) \bullet \hat{c}(n-l)^* + c(n) \bullet v_{cpich}(n-l) + v_{cpich}(n) \bullet v_{cpich}(n-l)\}$$

Since noise samples $v_{cpich}(n)$ taken at any lag $l \neq 0$ are uncorrelated, their means are null. Hence,

$$\phi_{cc}(l) = \mathrm{E}\{c(n)\hat{c}(n-l)^*\}$$

4) In a fourth step, we estimate said complex factor β with said estimates in step 3.

We assume in a first instance that the DPCH and CPICH channels have the same symbol rate 1/T, i.e. the same spreading factor *M*. We shall extend the treatment to the more general case of different spreading factors successively.

Consider the instantaneous ML channel estimates provided by (8). Then, under the assumption of modeling all the interference and noise terms as an AWGN, the complex factor β can be estimated as follows.

$$\hat{\beta}_l = \frac{E\{\hat{c}_{dpch}(n)\hat{c}^*_{cpich}(n-l)\}}{E\{\hat{c}_{cpich}(n)\hat{c}^*_{cpich}(n-l)\}} = \frac{\phi^*_{dc}(l)}{\phi^*_{cc}(l)} = \frac{r_{dc}(l)}{r_{cc}(l)} = \beta \qquad \forall l \neq 0 \quad (11)$$

Since again with the AWGN noise model, noise samples taken at any lag $l \neq 0$ are uncorrelated, the mean of such noise samples are equal to 0. Hence, one advantage is that we get rid of the noise and interference that annoy us on the CPICH and DPCH channels.

In practice the true correlations $r_{dc}(l)$ and $r_{cc}(l)$ can't be calculated, as a mean has to be calculated on an infinite number of samples. Therefore, these correlations need to be replaced by correlation estimates $\hat{r}_{dc}(l)$ and $\hat{r}_{cc}(l)$.

In a preferred non-limitative embodiment, we use sample moments as $\hat{r}_{dc}(l)$ and $\hat{r}_{cc}(l)$, i.e.,

$$\hat{r}_{dc}(l) = \frac{1}{N_p}\sum_{n=0}^{N_p-1}\hat{c}_{dpch}(n)\hat{c}^*_{cpich}(n-l) \quad (12)$$

and

$$\hat{r}_{cc}(l) = \frac{1}{N_p}\sum_{n=0}^{N_p-1}\hat{c}_{cpich}(n)\hat{c}^*_{cpich}(n-l) \quad (13)$$

so that

$$\hat{\beta}_l = \frac{\sum_{n=0}^{N_p-1}\hat{c}_{dpch}(n)\hat{c}^*_{cpich}(n-l)}{\sum_{n=0}^{N_p-1}\hat{c}_{cpich}(n)\hat{c}^*_{cpich}(n-l)} \quad (14)$$

where N_p is the number of pilot symbols comprised in the second sub-channel DPCCH.

Note that if the quality of the estimation is not good enough because of a too low value of the number of symbols N_p, one may increase N_p by including DD channel

estimates, when reliable symbol decisions can be taken at the output of the RAKE receiver, that is to say when there is a good signal to noise ratio SNR.

In a preferred non-limitative embodiment, in order to improve the estimates, further noise rejection can be achieved by considering the weighed sum

$$\hat{\beta} = \sum_{l=1}^{L} \gamma_l \hat{\beta}_l \quad \text{, L being an integer} \qquad (15)$$

as an estimate for β, where the γ_l are weighed coefficients and are to be chosen in order to account for the reliability of the estimates $\hat{\beta}_l$. To derive the optimal weighed coefficients γ_l, one should determine the joint statistic of $\hat{\beta}_l$, but that cannot be done in analytical way, as it would be too complex.

But, one can notice that the quality of the estimates $\hat{\beta}_l$ is likely to decrease as the lag l increases. Indeed, when the lag l increases, any channel correlation decreases whereas the noise increases. Hence, for a fixed number of pilot symbols N_p the noise **bias** in the sample moments estimates (9) and (10) increases as a channel correlation decreases.

Thus, in order to limit the algorithm complexity, in a preferred embodiment, we can limit the estimator (15) to the first term, that is to say we regard the first term as having the optimal weighed coefficient $\gamma_1 = 1$. Indeed, it would be too complicated to make a weighted mean on the other terms $\hat{\beta}_l$, $l \neq 1$. Hence, we have:

$$\hat{\beta} = \frac{\sum_{n=0}^{N_p-1} \hat{c}_{dpch}(n)\hat{c}^*_{cpich}(n-1)}{\sum_{n=0}^{N_p-1} \hat{c}_{cpich}(n)\hat{c}^*_{cpich}(n-1)} \qquad (16)$$

5) In a fifth step, from steps 2 and 4, we build a first sub-channel DPDCH multipath (MP) coefficient estimate $\tilde{c}_{dpdch}(n) = \hat{\beta}\, \tilde{c}_{cpich}(n)$.

In order to build the first sub-channel DPDCH estimate, for all the multipath component MP, we repeat the steps 1 to 5.

General DPDCH Channel Estimate at different symbol rates from CPICH and DPCCH

As we have seen before, we have assumed in a first instance the first CPICH and second DPCH channels to have the same symbol rate 1/T. But, in reality, it is not always the case. When the second channel's DPCH spreading factor M is not equal to 256, the symbol rates of the second channel DPCH and of the first channel CPICH differ.

Hence, in a preferred non limitative-embodiment, the channel estimates of either the first channel CPICH or the second channel DPCH are interpolated to match the symbol rate of the other channel estimates in order to apply the previous algorithm (16) so as to estimate the complex factor β.

This step of interpolation is done after the first step of the ML estimates of the DPCH and CPICH $\hat{c}_{dpch}(n)$, $\hat{c}_{cpich}(n)$ described before.

In deriving the ML-static channel estimates (8), both first CPICH and second DPCH channels are implicitly assumed constant over the longer symbols period between said channels, i.e. between two $c(n)$ and $c(n+1)$.

Therefore, in a preferred non-limitative embodiment, it is easier to make an interpolation to the lower symbol rate 1/T between said first CPICH and second DPCH channels. In particular, when the DPCH spreading factor M is smaller than 256, the interpolation to the CPICH symbol rate is obtained by averaging DPCH ML static channel estimates over a number of symbols equal to $256/M$. Whereas, when M=512, the DPCH symbol rate being lower than the one of the CPICH, then the average is taken on the CPICH over two symbol channel estimates.

In another embodiment, interpolation can be done to the highest symbol rate 1/T between DPCH and CPICH.

One can notice that these embodiments have the advantage of not coloring the estimation noise i.e. it doesn't imply change on the estimate algorithm (16), as we have seen that noise samples are uncorrelated.

Estimator Properties

As we have seen, an estimator has been derived applying the well-known method of samples moments, where the cross- and auto-correlation were replaced by the corresponding sample estimates. Moreover, the least complexity form of the estimator reduces to computing the cross- and the auto-correlation at one lag l as in the algorithm

(16). Unfortunately, even in this case statistical properties of the estimator cannot be derived in analytical way. It means that from the algorithm (16), we can not derived the performances of the estimator such as the number of pilot symbols N_p we need to have to reach a precision of 1%, the average precision of said estimator.... However it is straightforward to see that the estimator is asymptotically unbiased. For $N_p \to \infty$, $\hat{r}_{dc}(l) \to r_{dc}(l)$ and $\hat{r}_{cc}(l) \to r_{cc}(l)$ so that $\hat{\beta} \to \beta$ for any correlation lag. It means that the estimates are almost closed to the true values, which is a sufficient performance. Note that the number of pilot symbols N_p depends on the SNR. The higher the SNR is, the lower the number of N_p is needed to reach a determine performance.

Estimator Architecture

Fig. 2 represents a schematic architecture for an implementation of a RAKE receiver RECEIV comprising a proposed channel estimator ESTIMOR using implementing the algorithm (16). The estimator generates an estimate of β every slot.

The RAKE receiver comprises:

- a descrambler DESCRAMB,
- a de-spreader CPICH_DESPREAD and DPCH_DESPREAD associated to each CPICH and DPCH channels respectively , in order to de-spread a received spreading signal,
- the estimator ESTIMOR, and
- a demultiplexor DEMUX in order to recover the two physical channels DPCCH and DPDCH.

Note that a process of modulation of a transmitted signal is made with a pseudo-random sequence in order to limit the interferences between different base stations BS, this process, well known by the person skilled in the art, being called scrambling. At the receiver side, a de-scrambling needs to be achieved.

The estimator ESTIMOR comprises:

- Multipath coefficient estimate means CPICH_ESTIM that are adapted to calculate the estimate $\tilde{c}_{cpich}(n)$ of the multipath coefficient of step 2) from the CPICH channel's pilot symbol $s_{cpich}^{*}(n)$,
- Factor estimate means BETA that are adapted to calculate the complex factor β,
- A first interpolator INTERP1 on the CPICH branch, which is enabled when the DPCH spreading factor is equal to 512 i.e. when the CPICH symbol rate is lower than the DPCH one's,

- A second interpolator INTERP2 on the DPCCH branch, which is always enabled but not when the DPCH spreading factor is equal to either 256 or 512, i.e. when the DPCH symbol rate is lower than the CPICH one's.
- Thus, when the spreading factor is equal to 256, none of the first or second interpolators are enabled, as it means that the CPICH and DPCH have the same symbol rate.

In practice, in a preferred embodiment, nearest neighbor interpolation between two samples will be applied by the first INTERP1 and second INTERP2 interpolators, as it is shown to be a proper solution for this purpose. Indeed, it is the least complexity interpolation method while it yields a signal distortion due to such interpolation significantly below the noise level for practical ranges of Signal Interference (MAI) Noise Ratio SINR.

In a non-limitative preferred embodiment, the estimator ESTIMOR further comprises:

- Low-pass filtering means LPF1 and LPF2 for each channel CPICH and DPCCH respectively.

The estimates of the cross- and auto-correlation (9) and (10) can be significantly improved by low-pass filtering successive estimates taken over several slot periods. Indeed, the beamforming weight vector $\boldsymbol{w}$ has a variation rate that depends on the variation rate of the path angles θ. In practice path angles θ, between the mobile terminal MT and the base stations BS, are likely to stay approximately constant over several slot periods i.e. there are only small angle variations. Slight beamforming adjustment may be needed although to track those small angle variations. Therefore β can be estimated and tracked by properly low-pass filtering successive estimates of the cross- and auto-correlation (9) and (10), e.g. with an exponential forgetting factor (which means that a smaller weighting is applied on the older estimates), improving the noise suppression in the estimate. Note that as commonly known, any low-pass filter comprises a memory MEM to track the estimates.

But, when the base station BS abruptly changes the beamforming weight vector $\boldsymbol{w}$, the low-pass filters can't be used and usually the base station BS has to send a signaling MEM_RESET to the mobile terminal MT that provides a reset of the low-pass filter memory MEM. In case, such a signaling between the base station BS and a mobile terminal MT is not possible, the low-pass filters have to be removed. But whenever the low-pass filters are reset or removed, no signaling is needed to inform the mobile terminal MT that transmit beamforming is used or not. Hence, said mobile terminal MT can keep estimating the factor β independently of the presence of transmit beamforming, and when no transmit

beamforming is performed at the base station BS, β will just be a real factor accounting for the transmit power offset between the CPICH and DPCH channels.

In a non-limited preferred embodiment, the estimator ESTIMOR further comprises:

- Two delay lines Z^{-L} and $Z^{-L(256/M)}$ for the compensation of the delays introduced by the low-pass filters LPF, the delay being known and equal to L CPICH symbol periods.

One can notice that the algorithm (16) required operations to generate an estimate of β every slot, which are $2N_p$ products, $2(N_p-1)$ sums (N_p being the number of pilot symbols), and 1 division in addition to the operations required to low-pass filter the cross- and auto-correlation estimates. In the case of the use of low-pass filters, the least complexity solution for this purpose consists of adopting, in a preferred embodiment, a first order IIR KALMAN filter (with a suitable loop gain) as low pass filter, said filter being well-known by the person skill in the art. The additional operations with this solution would be in general four products and two sums (i.e., two products and one sum per low-pass filter).

In a non-limitative preferred embodiment, the estimator ESTIMOR further comprises:

- A third interpolator INTERP3, which is enabled when the spreading factor of the DPCH channel is lower than 256, and which is adapted to interpolate one of the symbol rate of the estimate $\hat{\beta}\,\tilde{c}_{cpich}(n)$ calculated at step 5) or of the symbol rate of the useful data of the DPDCH channel to match the other one.
- A multiplier that multiplies the DPDCH multipath MP coefficient estimate to the data of said DPDCH channel, the output being sent to a combiner RECEIV_COMB of the RAKE receiver.

Note that the structure in Fig. 2 is to be replicated for all RAKE receiver fingers. Each structure like the one in Fig. 2 has an output that is sent to the combiner RECEIV_COMB of the RAKE receiver in order to combine all the paths of a beam together.

Conclusions

Before the optimization of a receiver, there is a phase of acquisition through a synchronization channel SCH, well known by the person of ordinary skill in the art, which consists in an estimate of the delays τ_p and the signal energy of the paths.

According to the method of the invention described above, we have implicitly assumed that when the receiver can detect a path and estimate the associated delay on the CPICH through the synchronization channel SCH, a corresponding path at the same delay exists on the DPCH even though with a different complex amplitude. As we have seen, this assumption allowed the derivation of a DPCH channel estimation algorithm with low complexity, good performances and high flexibility, able to cope with the problem of channel estimation in the presence of transmit beamforming.

It will be noted that under certain circumstances a path carrying significant signal energy on the CPICH can be detected at a certain delay by the synchronization channel SCH, while the corresponding DPCH path signal energy is too weak to be detected. Conversely, it may happen that a path that carries significant signal energy on the DPCH is actually undetectable on the CPICH, the acquisition phase being mute. The first case would lead to an estimate of β with a very small magnitude. Thus, the algorithm described above would still work properly and the RAKE receiver would simply drop the corresponding finger on the DPCH. The second case is much more critical because there would be no way to exploit the information provided by the SCH and by the CPICH. Under these circumstances the estimator (13) cannot operate correctly and one should search for paths by directly using the second sub-channel DPCCH estimate, which is rather complex as it has been above-mentioned.

It is to be understood that the present invention is not limited to the aforementioned embodiments and variations and modifications may be made without departing from the spirit and scope of the invention as defined in the appended claims. In the respect, the following closing remarks are made.

It is to be understood that the present invention is not limited to the aforementioned UMTS application. It can be use within any application of DS-CDMA using a receiver.

It is to be understood that the method according to the present invention is not limited to the aforementioned implementation.

There are numerous ways of implementing functions of the method according to the invention by means of items of hardware or software, or both, provided that a single item of hardware or software can carries out several functions. It does not exclude that an assembly of items of hardware or software or both carry out a function. For example, the estimates of step 3) can be combined with the estimate of step 4), thus forming a single function without modifying the method of channel estimate with transmits beamforming in accordance with the invention.

Said hardware or software items can be implemented in several manners, such as by means of wired electronic circuits or by means of an integrated circuit that is suitable programmed respectively. The integrated circuit can be contained in a computer or in the receiver. In the second case, the receiver comprises means of the receiver adapted to make the de-scrambling, de-spreading, ..., said means being hardware or software items as above stated. In the same way, the estimator within the receiver, comprises maximum likelihood estimate means adapted to make the ML estimates of step 1), first multipath coefficient estimate means adapted to make the multipath coefficient estimate of step 2), correlation estimate means adapted to make the correlations of step 3), complex factor estimate means that are adapted to make the factor complex estimate of step 4), and second multipath coefficient estimate means that are adapted to make the first sub-channel multipath coefficient estimate of step 5), as described previously, said means being hardware or software items as above stated. Of course, as mentioned above, said means can be an assembly of items of items of hardware or software or both carry out a function, or a single item carrying out several functions.

The integrated circuit comprises a set of instructions. Thus, said set of instructions contained, for example, in a computer programming memory or in a receiver memory may cause the computer or the receiver to carry out the different steps of the estimate method.

The set of instructions may be loaded into the programming memory by reading a data carrier such as, for example, a disk. A service provider can also make the set of instructions available via a communication network such as, for example, the Internet.

Any reference sign in the following claims should not be construed as limiting the claim. It will be obvious that the use of the verb "to comprise" and its conjugations do not exclude the presence of any other steps or elements besides those defined in any claim. The word "a" or "an" preceding an element or step does not exclude the presence of a plurality of such elements or steps.

CLAIMS

1. A method for channel estimate with transmit beamforming, said channel comprising a first (CPICH) and second (DPCH) channels, said second channel (DPCH) comprising a first sub-channel (DPDCH) and a second sub-channel (DPCCH), characterized in that it comprises the steps of:

- Model said first (CPICH) and second (DPCH) channels, said models being based on multipath (MP) components, a multipath component being characterized by a time-varying complex coefficient ($c_p(t)$) and a delay (τ_p), said time-varying complex coefficient having a first channel (CPICH) multipath coefficient ($c_{pcpich}(t)$), and a second channel (DPCH) multipath coefficient ($c_{pdpch}(t) = \beta_p c_{pcpich}(t)$) depending on a complex factor (β_p) due to the transmit beamforming, each *p*-th multipath component being processed as follows :
- 1) Building a second channel (DPCH) and first channel (CPICH) instantaneous maximum likelihood (ML) channel multipath (MP) coefficients estimates ($\hat{c}_{dpch}(n)$, $\hat{c}_{cpich}(n)$);
- 2) Building an estimate ($\tilde{c}_{cpich}(n)$) of the first channel (CPICH) multipath coefficient;
- 3) Building estimates of a second channel (DPCH) and first channel (CPICH) cross-correlation ($\mathrm{E}\{\hat{c}_{dpch}(t)\hat{c}_{cpich}(n-l)^*\}$) and of a first channel (CPICH) auto-correlation ($\mathrm{E}\{\hat{c}_{cpich}(n)\hat{c}_{cpich}(n-l)^*\}$) from maximum likelihood (ML) estimates of step 2 at non-zero lag for noise suppression ($\tau \neq 0$);
- 4) Estimate said complex factor (β_p) with said estimates in step 3;
- 5) From steps 1 and 4 build a first physical channel (DPDCH) *p*-th multipath (MP) coefficient estimate ($\tilde{c}_{dpdch}(n) = \hat{\beta}\ \tilde{c}_{cpich}(n)$).

2. A method as claimed in claim 1, characterized in that the second channel (DPCH) maximum likelihood (ML) channel multipath (MP) coefficient estimate ($\hat{c}_{dpch}(n)$) is only based on a maximum likelihood (ML) channel multipath (MP) coefficient estimate of the second sub-channel (DPCCH).

3. A method as claimed in claim 1, characterized in that the estimate of a second channel (DPCH) and first channel (CPICH) cross-correlation ($E\{\hat{c}_{dpch}(n)\hat{c}_{cpich}(n-l)^*\}$) is an estimate of the second sub-channel (DPCCH) and first channel (CPICH) cross-correlation.

4. A method as claimed in claims 1, characterized in that one of the instantaneous maximum likelihood (ML) channel multipath (MP) coefficients estimates of step 2 is interpolated to match a symbol rate of the other channel estimate.

5. A method as claimed in claim 4, characterized in that an interpolation is done to the lower symbol rate between the first (CPICH) and second (DPCH) channels.

6. A method as claimed in claim 1, characterized in that the second channel (DPCH) and first channel (CPICH) cross-correlation's estimate and the first channel (CPICH) auto-correlation's estimate are low-pass filtered.

7. A method as claimed in claim 1, characterized in that a first channel (CPICH) multipath coefficient's estimate is done by a linear prediction.

8. A method as claimed in claim 1, characterized in that an estimate of said complex factor (β) is achieved with a method of sample moments ($\hat{r}_{dc}(k)$ and $\hat{r}_{cc}(k)$).

9. A method as claimed in claim 1 or 8, characterized in that an estimate of said complex factor (β) is a weighed sum of estimates ($\hat{\beta} = \sum_{k=1}^{K} \gamma_k \hat{\beta}_k$) at lag l.

10. A method as claimed in claim 9, characterized in that an estimate of said complex factor (β) is limited to the first term of said weighed sum.

11. A receiver utilizing said method as claimed in claim 1.

12. A computer program product for a receiver, comprising a set of instructions, which, when loaded into said receiver, causes the receiver to carry out the method claimed in claims 1 to 10.

13. A computer program product for a computer, comprising a set of instructions, which, when loaded into said computer, causes the computer to carry out the method claimed in claims 1 to 10.

14. An estimator for channel estimate with transmit beamforming, said channel comprising a first (CPICH) and second (DPCH) channels, said second channel (DPCH) comprising a first sub-channel (DPDCH) and a second sub-channel (DPCCH), characterized in that it comprises:

- 1) Maximum likelihood estimate means that are adapted to calculate a second channel (DPCH) and first channel (CPICH) instantaneous maximum likelihood (ML) channel multipath (MP) coefficients estimates ($\hat{c}_{dpch}(n)$, $\hat{c}_{cpich}(n)$);
- 2) First multipath coefficient estimate means (CPICH_ESTIM) that are adapted to calculate an estimate ($\tilde{c}_{cpich}(n)$) of the first channel (CPICH) multipath coefficient;
- 3) Correlations estimate means that are adapted to calculate a second channel (DPCH) and first channel (CPICH) cross-correlation ($\mathrm{E}\{\hat{c}_{dpch}(t)\hat{c}_{cpich}(n-l)^{*}\}$) estimate and a first channel (CPICH) auto-correlation ($\mathrm{E}\{\hat{c}_{cpich}(n)\hat{c}_{cpich}(n-l)^{*}\}$) estimate from said maximum likelihood (ML) estimates ($\hat{c}_{dpch}(n)$, $\hat{c}_{cpich}(n)$);
- 4) Complex factor estimate means that are adapted to calculate a complex factor (β) from said estimates of cross-correlation and auto-correlation;
- 5) Second multipath coefficient estimate means that are adapted t calculate a first sub-channel (DPDCH) p-th multipath (MP) coefficient estimate ($\tilde{c}_{dpdch}(n) = \hat{\beta}\,\tilde{c}_{cpich}(n)$) from the complex factor (β) and the estimate ($\tilde{c}_{cpich}(n)$) of the first channel (CPICH) multipath coefficient.

15. A receiver comprising said estimator as claimed in claim 14.

Method for channel estimate with transmit beamforming

ABSTRACT

The present invention relates to a method for channel estimate with transmit beamforming, said channel comprising a first (CPICH) and second (DPCH) channels, said second channel (DPCH) comprising a first sub-channel (DPDCH) and a second sub-channel (DPCCH). It is characterized in that it comprises the steps of:

- Model said first (CPICH) and second (DPCH) channels, said models being based on multipath components, a multipath component being characterized by a first channel (CPICH) multipath coefficient, and a second channel (DPCH) multipath coefficient depending on a complex factor (β_p) due to the transmit beamforming, each multipath component being processed as follows :
- 1) Building a second channel (DPCH) and first channel (CPICH) instantaneous maximum likelihood channel multipath coefficients estimates;
- 2) Building an estimate ($\tilde{c}_{cpich}(n)$) of the first channel (CPICH) multipath coefficient;
- 3) Building estimates of a second channel (DPCH) and first channel (CPICH) cross-correlation and of a first channel (CPICH) auto-correlation from step 2 at non-zero lag;
- 4) Estimate said complex factor (β) from step 3;
- 5) From steps 1 and 4 build a first physical channel (DPDCH) p-th multipath coefficient estimate.

Use: UMTS Receiver

Reference: Fig. 2

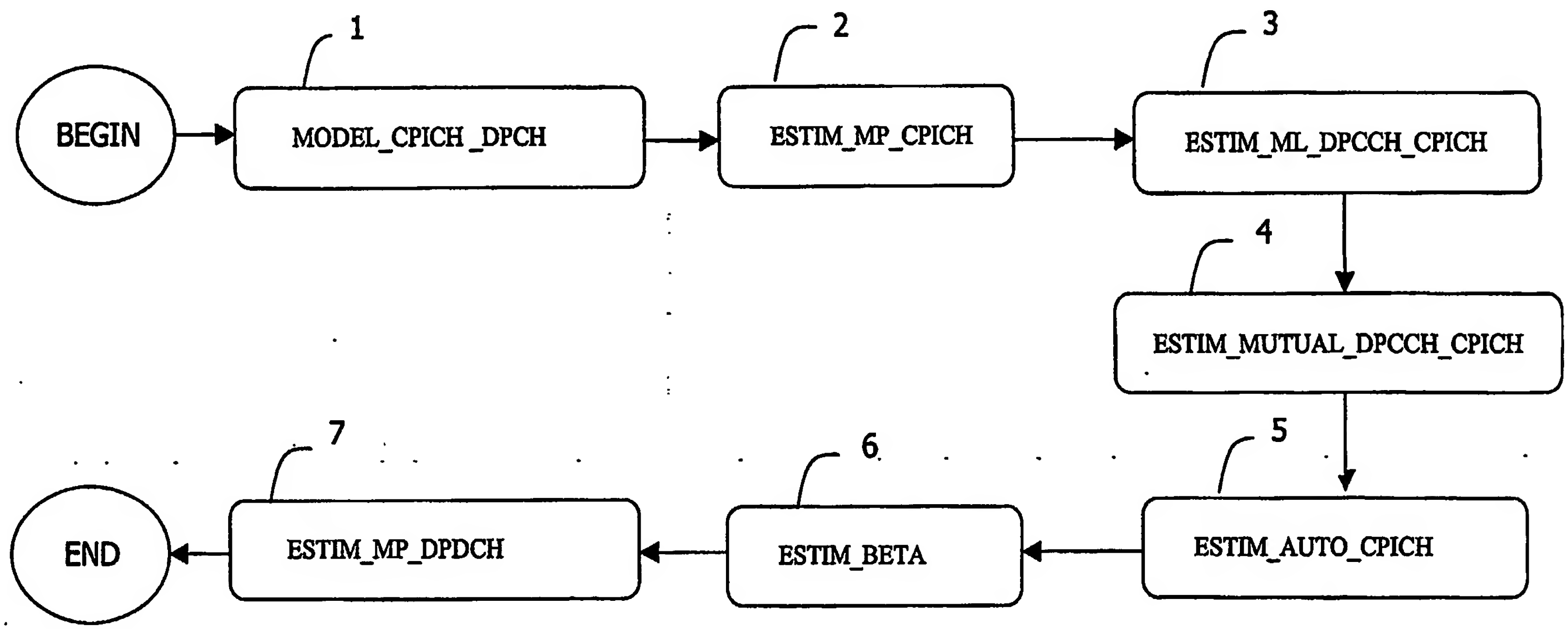
BEGIN
1
MODEL_CPICH_DPCH
2
ESTIM_MP_CPICH
3
ESTIM_ML_DPCCH_CPICH
4
ESTIM_MUTUAL_DPCCH_CPICH
5
ESTIM_AUTO_CPICH
6
ESTIM_BETA
7
ESTIM_MP_DPDCH
END

FIG. 1

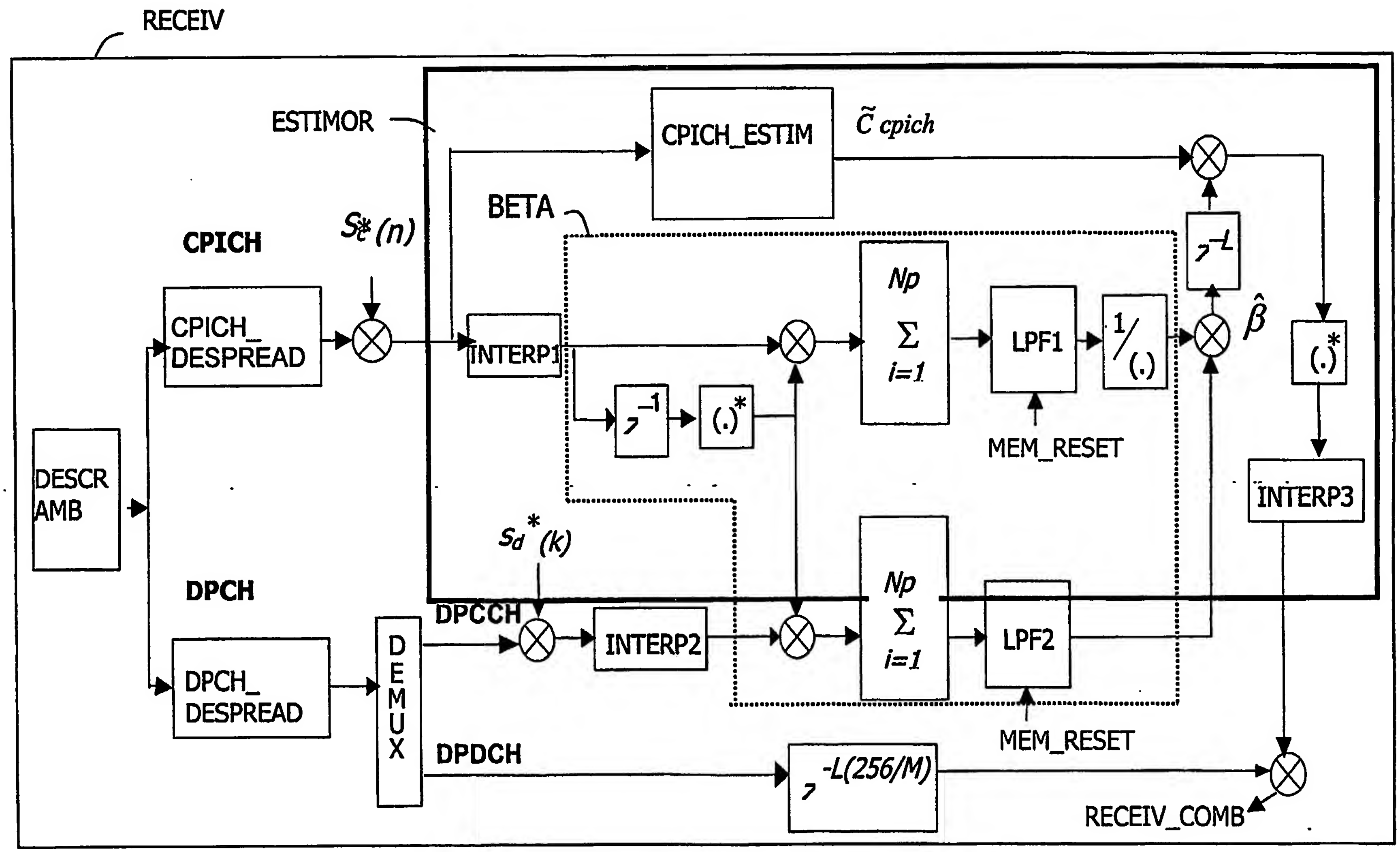
RECEIV
ESTIMOR
CPICH_ESTIM
C̃ cpich
BETA
CPICH
Sc*(n)
CPICH_
DESPREAD
INTERP1
Np
Σ
i=1
LPF1
1/(.)
β̂
(.)*
MEM_RESET
INTERP3
DESCR
AMB
DPCH
sd*(k)
DPCCH
INTERP2
LPF2
DPCH_
DESPREAD
DEMUX
DPDCH
MEM_RESET
RECEIV_COMB

FIG.2